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Long range television reception.

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LONG RANGE TELEVISION RECEPTION

BY

J. W. NEWLAND, JR.

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-

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by

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PREFACE

The investigation of the long range television reception problem was commenced in the fall of 1950 at the United States Naval Postgraduate School and was continued through the winter. This work was done in order to give the writer background material for receiving distant television stations in the vicinity of Albany, New York.

The writer wishes to express his appreciation for general guidance to Associate Professors P. E. Cooper and Professor C. E. Menneken of the United States Postgraduate School.

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TABLE OF SYMBOLS AND ABBREVIATIONS

A = voltage gain

C_{IN} = tube input capacitance

C_{OUT} = tube output capacitance

Δf = bandwidth, in cycles/seconds

E = rms value of thermal-noise voltage

E_{ANT} = antenna signal voltage

E_b = average plate voltage

E_{c2} = average screen grid voltage

$E_{g\,NOISE}$ = noise voltage at control grid

$E_{g\,sig}$ = signal voltage at control grid

$E_{p\,NOISE}$ = noise voltage at plate

$E_{p\,sig}$ = signal voltage at plate

F = noise figure

G_F = available power gain of a network for a particular frequency

g_m = tube grid-plate transconductance

I_a = average cathode current

I_b = average plate current

I_{c2} = average screen current

k = Boltzmann's constant = 1.37×10^{-23} watt-second/degree

μ = tube amplication factor

N = available noise power

R = resistance component of impedance, in ohms

R_{ANT} = characteristic resistance of the transmission line from the antenna

TABLE OF SYMBOLS AND ABBREVIATIONS (CONT.)

R_{eq} = noise equivalent resistance

R_g = antiresonant resistance of the grid circuit

R_{in} = equivalent resistance of the input circuit

R_p = internal tube resistance

R_s = equivalent resistance of the input signal noise voltage

R_T = equivalent noise resistance of the tube

S = output available signal

S_g = input available signal

T = absolute temperature

T_1 = noiseless tube equivalent

CHAPTER I

INTRODUCTION

Throughout the United States today millions of people relax in their homes enjoying the pleasure of the modern electronic miracle of television. There are also many more millions who are deprived of this enjoyment entirely, or are limited to the reception of only one or two stations. This may be due to the fact that in a sparsely populated area the normal television range cannot support the expense of a station, or in other areas there are insufficient stations due to the television "freeze". It would seem reasonable to assume at this time, that for the most part these people now deprived of television will be unable to obtain this enjoyment in the future even with the allocation of new frequency bands.

The situation as it now exists is very similar to that of the early days of radio when people were happy and content if they could receive a single station. At that time there was sufficient space in the frequency spectrum and on their radio dial for hundreds of additional stations and today although a person can receive but one television station his receiver is ready and will receive additional stations if the signal were present at the input terminals. It is quickly apparent that the reception of additional stations on a television receiver increase the entertainment value of the receiver beyond measure.

The approach which will be taken to the problem of receiving television in a deprived area will be to accept the situation as it exists today, and which will be present to a great degree in the future, and attempt to show the controlling factors in the design of a long

range fringe area television antenna system. By means of this fringe area antenna system the signals from distant transmitting stations will be fed to the input terminals of the present day receiver at a much greater strength than they would normally be received increasing the reception range to a very marked extent.

The basic receiving system whose design factors will be discussed will include antenna, matching and coupling networks, and pre-amplifier. Since the most important factor in the long range reception of television is the magnitude of the input circuit noise generated by the antenna, coupling circuits and the first tubes, the noise problem will be discussed in considerable detail. Such factors as tube types, circuitry, and an approach to the noise problem for any tube circuit will be covered in regards to the noise problem in the pre-amplifier. Types of antennas, matching and coupling circuits will also be discussed in relation to the overall receiving system.

CHAPTER II

TYPES OF NOISE

There are countless causes of noise in electronic equipment. Even though there are many causes, the more important ones may be classified as random and non-random noise.

Random noise is that noise due to the atomic nature of matter and electricity. It may be said that this random noise is but one manifestation of the fluctuations that occur on a microscopic scale in all forms of matter and energy in which it is found that the molecules, ions, electrons and even photons are in a perpetual state of more or less random motion. The simplest and earliest known example of this state of chaotic motion underlying all matter may be observed, with the aid of a microscope, in a clear liquid containing a suspension of very fine particles. By this simple experiment in "Browman movement" the existence and movement of molecules are proved directly. This fundamental type of noise present in electronic equipment is remarkable in that it is completely without regularity in its detailed properties and has the important wave-shape property of the addition of noise powers.

Non-random noise may be simply defined as noise that is not of a random nature.

A difference in the two types of noise which will appear in later chapters is the manner in which their wave shapes vary with bandwidth. In the case of random noise, if the bandwidth is increased, the general character of the signal remains the same, showing only an

increase in average height and a finer structure. On the other hand, the effect of increased bandwidth on the wave form of non-random noise is to bring out the detail of the wave form. The effect of increased bandwidth in this case is similar to bringing the wave shape of the non-random noise into sharper focus.

Of the various kinds of noise, the simplest and perhaps also the most important is that called "thermal noise". This random type of noise was discovered in 1928 by J. B. Johson and at the same time H. Nyquist was able to show that the thermal-noise voltage generated in an impedance Z is given by the equation

$$E^2 = 4RkT\Delta f$$

where E = rms value of thermal-noise voltage

R = resistive component of the impedance, in ohms

T = absolute temperature

k = Boltzmann's constant = 1.37×10^{-23} watt-second/deg

Δf = bandwidth, in cycles/sec

The above equation shows that thermal noise has a uniform distribution of power throughout the frequency spectrum and that the thermal-noise voltage generated depends only on the resistive component of Z and is independent of the reactive component.

The other most important kind of noise is that called "shot noise". This random type of tube noise was discovered by W. Schottky in 1918 and consists of the combined effect of a large number of independently cathode emitted electrons. The emission current is never steady but exhibits minute fluctuations due to the finite charge of an electron in combination with its random emission.

A source of noise, especially in warm weather and warm climates, is the noise of atmospheric origin called "static" in the United States. Much of this noise originates in thunderstorms, in discharges between clouds and in similar ways. Atmospherics are of the general characteristic of random noise but their intensity is spasmodic. The field strength of static appears on the average to be approximately inversely proportional to frequency. Above 50 mc, practically all atmospherics are local, for there is usually no reflection from the Heaviside layer.

A number of sources of random noise although of minor nature compared to thermal and shot noise are as follows:

Flicker Effect - noise superimposed on shot noise due to the emission surface of cathode changing.

Contact and Breakdown Noise - noise due to the breakdown of insulation or the loss of contact in minute paths in equipment components.

Dirt and grain-size Noise - noise due to minute irregularities in equipment's structure.

Incidental Tube Noise - noise due such causes as secondary emission collision ionization and fluctuations in emission of positive ions.

Interstellar Interference - noise of a quasi-random nature which appears to originate in the general region of the Milky Way.

Thus far we have considered only random noise. However, there is also noise in electronic equipment that is not of a random nature and that has characteristic wave shapes of its own. This includes such

descriptive noise kinds as "hum", "howl", "wow", "motorboating", "hiss", "scratch", "flutter wow", etc. All of these types of non-random noise may be eliminated if proper precautions are taken.

CHAPTER III

DEFINITIONS

In order to be able to discuss the noise problem in a simple and intelligent manner, it is desirable to have a sound basis upon which to build. This can only be brought about by definition. The following, although uninteresting in many respects, probably contains the most important material found in this paper. Its importance may be judged by the fact that the definitions and ideas are the result of many years of work by the leaders in the field of noise and are the accepted basis for noise work as it stands today.

One of the basic definitions is that of "available power". A generator of real internal impedance R ohms and open circuit voltage E can deliver a maximum power of $E\ /4R$ watts into a resistive load. This maximum power which is available is by definition the "available power". This maximum power is actually delivered only when the generator is matched. Thus the "available power" is dependent only upon the internal impedance of the generator and no matter what load is connected to the generator is always $E\ /4R$ where R is the internal resistance in ohms of the generator.

From this definition of "available power" we come to "available noise power". It has been shown by Nyquist's work that any resistance of R ohms produces a noise voltage

$$E^2 = 4RkT\Delta f$$

where R = resistance in ohms
T - absolute temperature
R = Boltzmann's constant, 1.38 x 10
Δf = band width, in cycles/sec

Thus the resistance may be represented as a generator having an internal noise-free resistance of R ohms, and an open circuit r.m.s. voltage E. The available power from this generator is

$$\frac{E^2}{4R} = \frac{4kTR\Delta f}{4R} = kT\Delta f \text{ (available noise power)}$$

This "available power" is the "available noise power" delivered by the resistor to a noise free resistor of equal resistance.

In like manner the "available signal power" is the maximum power which can be removed from a signal generator.

If for any four terminal network the internal impedance of the output terminals is R and the signal voltage generated in the output circuit is E, then the available signal power in the output circuit is $E^2/4R$.

The "available power gain" of a four terminal network is defined as:

$$G = \frac{\text{available signal power in output circuit}}{\text{available signal power in the source feeding the input circuit}}$$

If G_f is the available power gain of a network for a particular frequency f, the effective bandwidth Δf of the network is

$$\Delta f = \frac{1}{G} \int G_f \, df$$

Friss defines the noise figure F of a network as the ratio of the "input" available signal to available noise power ratio" to "output available signal to available noise power ratio"

$$F = \frac{S_g / kT\Delta f}{S/N}$$

The noise figure F is not a measure of the excellence of the input signal-to-noise ratio, but merely a measure of the degradation suffered by the signal-to-noise ratio as the signal and noise passes through the

network in question. If the input signal and noise are extremely large compared to the amplifier noise, relatively little degradation is suffered, and the noise figure is good.

CHAPTER IV

NOISE-EQUIVALENT RESISTANCE

One of the greatest steps forward in the solution of the noise evaluation problem in vacuum tubes circuits was made by the trio of B. J. Thompson, D. O. North and W. A. Harris in 1941. Such results of their work as are applicable to the design of low noise r.f. amplifiers will be described as concisely as possible in order that the tremendous detail resulting from the numberous sources of noise in a vacuum tube circuit will not obscure the method of solution to be given in the following chapter.

Thompson, North and Harris approached the problem in this manner. For a single stage of amplification, as the input signal passes through the amplifier tube, noise of an unknown amount is added to the noise already present with the signal. The same results would be achieved if the actual tube were replaced by a theoretical tube having all of the characteristics exactly as that of the actual tube except for the fact the theoretical tube would be free of all sources of noise. By adding a noise generator to the grid circuit of the theoretical noise free tube such that the noise output of this tube was exactly the same as the actual tube the tube circuits would then be identical.

What Thompson, North and Harris did, was to evolve formulas for the output of the noise generator at the grid of the theoretical noise free tube employing specific operating conditions, and characteristics for various tube types.

They went one step further in evolving these formulas. By use of the Nyquist equation they transformed this random noise voltage into

what is called the noise-equivalent resistance of the tube. Take a triode for instance, the noise-equivalent resistance of a triode is,

$$R_{eq} = \frac{2.5}{g_m}$$

Where R_{eq} = noise equivalent resistance

g_m = grid-plate transconductance

Thus if a resistance is added to grid of a theoretical noise free triode operating with a grid-plate transconductance specified by the tube operating conditions, this resistance will generate a random noise voltage such that the noise output of the actual tube and the theoretical tube will be identical.

Use of the method of noise analysis described in following sections depends on assignment of noise-equivalent-resistance values to the "noise generators" and these values may be found from:

For triode amplifiers

$$R_{Eq} = \frac{2.5}{g_m}$$

For pentode amplifiers

$$R_{Eq} = \frac{I_b}{I_b + I_{c2}}\left(\frac{2.5}{g_m} + \frac{20 I_{c2}}{g_m^2}\right)$$

For triode mixers

$$R_{Eq} = \frac{4}{g_c}$$

or

$$R_{Eq} = \frac{2.5\,\bar{g}_m}{g_c^2}$$

For pentode mixers

$$R_{Eq} = \frac{I_b}{I_b + I_{c2}}\left(\frac{4}{g_c} + \frac{20 I_{c2}}{g_c^2}\right)$$

or

$$R_{Eq} = \frac{I_b}{I_b + I_{c2}}\left(\frac{2.5\,\bar{g}_m}{g_c^2} + \frac{20 I_{c2}}{g_c^2}\right)$$

For multigrid converters and mixers

$$R_{Eq} = \frac{20 I_b (I_a - I_b)}{I_a g_c^2}$$

where, R_{Eq} = noise-equivalent resistance

g_m = grid-plate transconductance

$\bar{g}_m$ = average transconductance (frequency converters and mixers)

$\bar{g}_c$ = conversion transconductance (frequency converters and mixers)

I_b = average plate current

I_{c2} = average screen-grid current

I_a = average cathode current

The following approximate reactions for triode and pentode mixers are useful when data required is not available.

As converter		As amplifier*
g_c	=	$\frac{1}{4} g_m$
I_b	=	$\frac{1}{4} I_b$
I_{c2}	=	$\frac{1}{4} I_{c2}$
R_{Eq}	=	$4 R_{Eq}$ **

* The values "as amplifier" refer to conditions at the peak of the assumed oscillator cycle.

** Pentode mixers only

There are certain restrictions which must be kept in mind when the above formulas are employed. In the case of triodes, the concept is applicable only in the absence of transit time effects and feedback. Additional noise sources are necessary to represent the noise behavior if either or both of the above phenomena are present. When tetrode or pentodes are used, the use of a single equivalent noise source additionally requires that the screen and suppressor be returned directly to the

cathode. It must also be realized that the use of a single noise source to represent a complex noise source implies some rather drastic assumptions concerning the frequency behavior of the impedances associated with the actual noise sources. Despite all of the assumptions, however, the concept leads to useful results.

CHAPTER V

TUBE TYPES

The equations given in the previous section give the theoretical expressions for the equivalent noise resistance of a tube, which is applicable either to a triode, or to a pentode the screen of which is connected to its cathode. These equations have been utilized in setting up the following table in which values of equivalent noise resistance and transconductances of common receiving tube are given. These values are calculated from the nominal values of the transconductance and plate and screen currents given in the RCA Tube Handbook. Average values of input and output capacitances, which are valid for the case when the tube has its cathode grounded, are given also; these values do not include socket or wiring capacitances.

It must be emphasized that these are theoretical values and although agreement between theory and experiment is quite good for some tube types, it is much less satisfactory for others. Another point to be remembered is that there is wide variation in the equivalent noise resistances of individual tube of a given type.

Tube	g_m μmhos	R_{Eq} ohms	C_{IN} $\mu\mu$f	C_{OUT} $\mu\mu$f
		Triode Amplifiers		
6AC7	11,250	220	11.0	4.0
6AK5	6,670	385	4.0	2.0
6C4	2,200	1,140	1.8	1.3
CF4	5,800	430	2.0	0.6
6J4	12,000	210	2.8	0.2
6J5	2,600	960	3.4	3.6
6J6	5,300	470	2.2	0.4
6SC7	1,325	1,890	2.2	0.3
6SL7	1,600	1,560	3.2	3.6
6SN7	2,600	960	2.9	1.0
7F8	5,650	440	2.8	1.4
9002	2,200	1,140	1.2	1.0
		Sharp Cut off Pentodes		
6AU6	4,500	2,550	5.5	5.0
6BC5	6,100	1,300	6.6	3.1
6CB6	6,200	1,470	6.3	1.9
		Remote Cut off Pentodes		
6BA6	4,400	3,550	5.5	5.5
6SK7	2,000	10,500	6.0	7.0

CHAPTER VI

METHOD

Any amplifier system which may be considered has three major sources of noise: noise which accompanies the input signal, thermal noise in the input circuit itself and tube noise. As has previously been mentioned there are also many other sources which enter the picture. These many sources leads us to the necessity in the design of our pre-amplifier and input circuit of finding a method by which all of these may be combined and the resultant effect foretold.

In evaluating any tube circuit with relation to thenoise from each source and for comparing the resultant noise level with signal level a good reference point is the control grid of the first tube. The key to the method which follows, lies in the important wave shape property of random noise that "noise powers add while noise voltages do not". From the Nyquist equation for the random voltage across the terminals of an impedance Z: $E^2 = 4RkT\Delta f$

and for two impedances in series:

$$E^2 = 4(R_1 + R_2)kT\Delta f$$

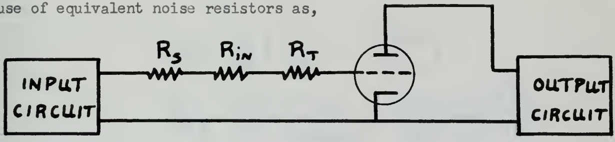
which means that a random noise voltages from any number of sources may be changed to their equivalent noise resistance and these independent resistances may be added together to give a total equivalent noise resistance. This total equivalent noise resistance may then be changed into the total noise voltage resulting from all sources.

Through the work of Harris and his associates an actual tube may be replaced by a theoretical noise free tube in whose grid circuit there is placed a fictitious resistor having a thermal noise voltage which produces the same noise voltage in the output of the theoretical tube as there is

in the output of the actual tube. This fictitious resistor, the grid equivalent noise resistor of the tube, is a legitimate and accurate equivalent because the noise energy of a resistor has essentially the same flat frequency spectrum as the tube noise.

As a result, the equivalent noise resistances of any source of voltage noise may be referred to the grid of the theoretical noise free tube and these resistances along with the grid equivalent noise resistor of the tube added together to give the total equivalent noise resistance in series with the grid circuit. This total equivalent noise resistor may then be transformed into the total equivalent noise voltage acting on the grid of the theoretical equivalent of the actual tube.

A single stage of amplification may be diagramed to illustrate the use of equivalent noise resistors as,



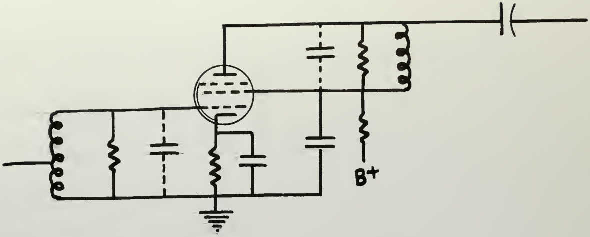
where R_S = equivalent resistance of the input signal noise voltage

R_{IN} = equivalent resistance of the input circuit

R_T = equivalent noise resistance of the tube

T_1 = noiseless tube equivalent

In order to illustrate completely the use of this method of attack on the noise problem in amplifier design, a pentode r.f. amplifier will be worked out in detail.



Tube: 6AK5

Typical operating conditions from the manual:

$$E_b = 120^v$$
$$E_{c2} = 120^v$$
$$g_m = 5000 \mu v$$
$$I_p = 7.5\ ma$$
$$I_{c2} = 2.5\ ma$$
$$I_K = 10\ ma$$

From North's work:
$$R_{Eq} = \frac{I_p}{I_K}\left(\frac{2.5}{g_m} + \frac{20 I_{c2}}{g_m^2}\right)$$

$$R_{Eq} = \frac{7.5}{10}\left(\frac{2.5}{5000\mu} + \frac{20 \cdot 2.5 \cdot 10^{-3}}{5000 \cdot 10^{-6}}\right)$$

$$R_{Eq} = 1500 \Omega$$

since
$$E^2 = 4RkT\Delta f$$

where, $T = 300^\circ$ absolute

$$\Delta f = 4\ mc$$
$$k = 1.374 \times 10^{-23}$$

$$E = 1.3 \times 10^{-10} \sqrt{R \Delta f}$$
$$E = 1.3 \times 10^{-10} \sqrt{1500 \cdot 4 \cdot 10^{6}}$$

$E = 10\ \mu v$ tube noise voltage referred to the grid.

Employing standard transformer theory to the input transformer we have,

$$\frac{E_{g\,sig}}{E_{ANT}} = \left(\frac{R_g}{R_{ANT}}\right)^{\frac{1}{2}}$$

where, $E_{g\,sig}$ = signal voltage on grid

E_{ANT} = antenna signal voltage

R_g = antiresonant resistance of the grid circuit

R_{ANT} = characteristic resistance of the transmission line from the antenna

Note from the above equation that for maximum signal on the grid, R_g must be as large as possible. By increasing R_g we increase the input signal and also the thermal noise due to R_g but this increase of both signal and noise voltage due to R_g also increases independent of tube noise voltage. This increase in R_g increases the S/N ratio and the output of the stage.

It has been shown by Bode (ref. 1, p. 213) that for a shunt resistance and capacitor, regardless of the location of band pass of a network in the frequency spectrum, the band pass is inversely proportional to the product of the resistor and capacitor. $\left[\Delta f \propto \frac{1}{RC}\right]$

For a parallel RLC circuit,

$$\Delta f = \frac{1}{2\pi RC}$$

for a 6AK5, input $C_{in} = 4\mu\mu f$

and wiring C plus $C_{in} = 7\mu\mu f$

assuming the band width we are designing for is 4MC

$$R_p = \frac{1}{2\pi C\Delta f}$$

$$R_p = \frac{1}{2\pi \cdot 7 \cdot 10^{-12} \cdot 4 \cdot 10^{6}} = 5700\,\Omega$$

which is the maximum shunt resistance.

If the antenna plus the lead-in transmission line combination is matched to the grid input circuit, the effective resistance shunting the grid would be halved. In order that the maximum shunt resistance be 5700 ohms the actual grid resistor must be twice this value or 11,400 ohms and must be matched to the antenna plus transmission line combination.

The noise voltage at the grid due to the input circuit will be.

$$E = 1.3 \times 10^{-10}\sqrt{R\Delta f}$$

$$E = 1.3 \times 10^{-10}\sqrt{5700 \cdot 4 \cdot 10^{6}}$$

$E = 20\,\mu V$ noise voltage at the grid due to input circuit

Since noise powers and not noise voltages add directly, the total noise voltage at the grid due to the tube and input circuit will be,

$$E = \sqrt{E_1^2 + E_2^2 \cdots}$$

$$E = \sqrt{20^2 + 10^2}$$

$$E = 22.4\,\mu V$$

If it is assumed that the antenna connected to the amplifier is a dipole with $300\,\Omega$ lead in and the signal voltage at the antenna terminals E_{ANT} is $20\,\mu V$, then from

$$\frac{E_{g\,sig}}{E_{ANT}} = \left(\frac{R_g}{R_{ANT}}\right)^{\frac{1}{2}}$$

and

$$E_{g\,sig} = \left(\frac{5700}{300}\right)^{\frac{1}{2}} E_{ANT}$$

$$E_{g\,sig} = 4.37 \cdot 20\,\mu V = 87.4\,\mu V$$

and at the input grid, $$S/N = \frac{87.4}{22.4} = 4.0$$

The next step in the noise analysis of the r.f. amplifier is to investigate what occurs beyond the grid of the first stage.

The output of the first stage is tuned and the wiring capacity between stages, plus the output capacity of the first stage, plus the input capacity of the second stage amounts to,

$$C = 14 \mu\mu f$$

$$\Delta f = 4 \text{ mc}$$

$$R = \frac{1}{2\pi C \Delta f}$$

$$R = \frac{1}{2\pi \cdot 4 \cdot 10^{6} \cdot 14 \cdot 10^{-12}}$$

$$R = 2,840$$

$$A = g_m R$$

$$A = 5000 \cdot 10^{-16} \cdot 2840$$

$$A = 14.2$$

$$E_{p\,sig} = A E_{g\,sig}$$

where $A =$

$E_{p\,sig} =$ signal voltage at plate

$E_{g\,sig} =$ signal voltage at grid

and since $E_{g\,sig} = 4.37 E_{ANT}$

$$E_{p\,sig} = 4.37 A E_{ANT}$$

$$E_{p\,sig} = 62 E_{ANT}$$

also $E_{p\,sig\,NOISE} = A E_{g\,NOISE}$

where, $E_{p\,NOISE} =$ noise voltage at plate

$E_{g\,NOISE} =$ noise voltage at grid

since $$E_{g\,NOISE} = 22.4\mu V$$

$$E_{p\,NOISE} = 14.2 \cdot 24 = 340\mu V$$

A convenient way in which to combine the interstage circuit thermal noise and the tube noise of the second stage, with the above noise voltage is to convert the noise voltage at the plate of the first tube to its equivalent noise resistance and add it to the interstage circuit resistance and the equivalent noise resistance of the second stage.

$$E = 1.3 \times 10^{-10}\sqrt{R\Delta f}$$

$$E_{p\,NOISE} = 1.3 \times 10^{-10}\sqrt{R_{p\,NOISE}\Delta f}$$

$$E_{p\,NOISE} = A\,E_{g\,NOISE}$$

$$A\,E_{g\,NOISE} = 1.3 \times 10^{-10}\sqrt{R_{p\,NOISE}\,\Delta f}$$

$$E_{g\,NOISE} = 1.3 \times 10^{-10}\sqrt{R_{g\,NOISE}\,\Delta f}$$

$$\frac{1.3 \times 10^{-10}\sqrt{R_{p\,NOISE}\,\Delta f}}{A} = 1.3 \times 10^{-10}\sqrt{R_{g\,NOISE}\,\Delta f}$$

$$\text{OR} \quad A^2 R_{g\,NOISE} = R_{p\,NOISE}$$

$$R_{g\,NOISE} = R_S + R_{IN} + R_T$$

disregarding the noise input from the signal in order that the circuit may be evaluated

$$R_{g\,NOISE} = R_{IN} + R_T$$

for first stage $$R_{g\,NOISE} = \frac{5700}{2} + 1500$$

$$R_{g\,NOISE} = 4350$$

$$R_{p\,NOISE} = A^2 R_{g\,NOISE}$$

$$= 14.2^2 \cdot 4350$$

$$= 870{,}000\,\Omega$$

It is now apparent that this value of equivalent noise resistance in

the input of the second stage is much greater than the equivalent noise resistance of the interstage circuit or the tube of the second stage and further investigation is not necessary from this standpoint since any further noise sources in the following circuitry would have only a minute effect.

Though approximations and assumptions have been made in the foregoing example it illustrates the basic procedure and indicates the simple, logical and powerful tool which is available in handling noise.

CHAPTER VII

CIRCUIT TYPES

In the previous section, a method for numerically finding the effect from noise on a stage of r.f. amplification along with its input circuit has been given. Now in order to compare the effect from noise of various tube configurations the idea of "noise figure" will be used. As has previously been stated, the "noise figure" F of a network is the ratio of the "input available signal to available noise power ratio" to "output available signal to available noise power ratio"

$$F = \frac{S_g / kT\Delta f}{S/N}$$

The important fact to remember is that the "noise figure" is merely a measure of the degradation suffered by the signal-to-noise ratio as the signal and noise passes through the network in question.

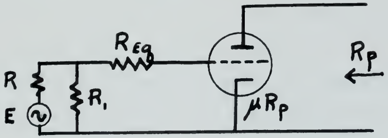
There are three basic circuits which may be employed in a stage a r.f. amplification,

(a) grounded cathode (conventional amplifier)

(b) grounded grid

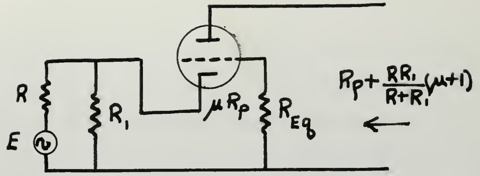
(c) grounded plate (cathode follower)

(a) grounded cathode amplifier



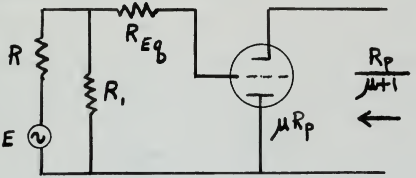
$$G = \frac{\mu^2 R}{R_p}\left(\frac{R_1}{R+R_1}\right)^2 \qquad F = \frac{R+R_1}{R_1} + \frac{R_{Eq}}{R}\left(\frac{R+R_1}{R_1}\right)^2$$

(b) grounded grid amplifier



$$G = \left(\frac{R_1}{R+R_1}\right)^2 (\mu+1)^2 \, R \Big/ \left[R_p + \frac{RR_1}{R+R_1}(\mu+1)\right] \qquad F = \frac{R+R_1}{R_1} + \frac{R_{Eq}}{R}\left(\frac{\mu}{\mu+1}\right)^2 \left(\frac{R+R_1}{R_1}\right)^2$$

(c) grounded plate



$$G = \left(\frac{R_1}{R+R_1}\right)^2 \frac{\mu^2 R}{(\mu+1)^2} \Big/ \frac{R_p}{\mu+1} \qquad F = \frac{R+R_1}{R} + \frac{R_{Eq}}{R}\left(\frac{R+R_1}{R_1}\right)^2$$

From the "noise figure" results of the three configurations it appears that the degradation suffered by the signal-to-noise ratio as the signal and noise passes through a stage of amplification is essentially independent of the manner in which the tube is connected.

Despite the fact that the noise-figure formulas are the same, the circuits cannot be considered as interchangeable from a noise standpoint, for reasons associated with the circuits to be used with the amplifier. Consider the problem of stability. It is well known that the cathode-

separation triode is prone to oscillate unless neutralized. For this reason the noisier pentode is generally preferred in grounded cathode amplifiers.

The largest consideration is the matching of the input source. With a long transmission line between antenna and pre-amplifier, the noise-figures for the grounded cathode and grounded plate amplifiers are identical. R_1 must be used to terminate the line and the noise figure for this termination becomes,

$$F_{TERM} = 2 + \frac{4 R_{EQ}}{R}$$

The limit of R being set by the type of transforming network and bandwidth.

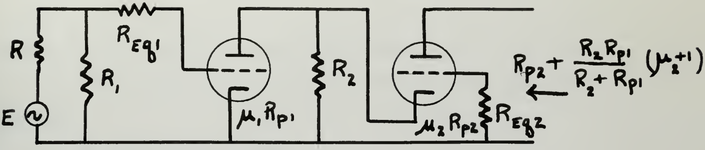
In the case of the grounded grid amplifier, however, the dynamic impedance may be arranged to terminate the line by choosing a tube with the proper g_m and plate load and the noise figure becomes

$$F_{TERM} = 1 + \frac{R_{EQ}}{R}$$

The cathode follower is not generally used because of its lack of voltage from grid to cathode. The impedance transformation from grid to output, however, does allow the use of a step-up circuit between the cathode and the input to the following stage. In this manner the noise figure of the second stage may be improved, and the voltage gain achieved.

The most widely used circuit, from the noise-figure point of view is the Wallman Amplifier (ref. 10). It is an almost incredible combination of favorable characteristics. It uses a grounded cathode triode amplifier driving a grounded-grid triode amplifier. The driving point impedance of the grounded grid amplifier, the load of the input amplifier, is so

low that there is little or no gain in the first stage and therefore is very stable. The good noise figure of the grounded grid stage is obtained from the high plate impedance of the first stage



Wallman amplifier

$$F_1 = \frac{R+R_1}{R_1} + \frac{R_{Eq1}}{R}\left(\frac{R+R_1}{R_1}\right)^2$$

$$G_1 = \frac{R\mu^2}{R_{p1}}\left(\frac{R_1}{R_1+R}\right)^2$$

$$F_2 = \frac{R_{p1}+R_2}{R_2} + \frac{R_{Eq2}\,\mu_2^2}{R_{p1}(\mu_2+1)^2}\left(\frac{R_{p1}+R_2}{R_2}\right)^2$$

$$F_{1,2} = 1 + \frac{R_{Eq1}}{R} + \frac{R_{Eq2}\,\mu_2^2}{R\mu_1^2(\mu_2+1)^2}$$

If μ_1 and μ_2 are large compared to unity as is usually the case, the noise figure of the combination is essentially that of the input triode alone, less the instability disadvattage of the grounded cathode amplifier without altering its noise characteristics.

CHAPTER VIII

COUPLING CIRCUITS

The purpose of the input coupling network is to transform the conductance of the signal source to a value that gives the optimum noise figure consistent with the bandpass characteristics of the amplifier.

Two general types of input networks appear the most logical and practical:

(1) the single tuned circuit

(2) the double tuned circuit

The single tuned circuit employs the fewest number of parts, occupies the least space, and is easy to align. The single tuned circuit however, because of its high Q does not always permit the bandwidth requirements to be met. In order to meet the bandwidth requirements, the coil must be loaded which in turn raises the noise figure. In some instances in long range television reception a narrowing of bandpass beyond the transmitted modulation frequency range while eliminating a portion of the intelligence, compensates for this by a far better signal to noise ratio and results in an improved picture. The alternative method of widening bandwidth is the use of a double-tuned input circuit.

The double tuned circuit has approximately double the bandwidth for the same transformed conductance as the single tuned circuit, and it has increased selectivity. The double tuned circuit has the disadvantages of increased size and complexity and greater dependence on circuit constants and in particular a greater dependence on tuning errors.

CHAPTER IX

ANTENNA

The satisfactory performance of any television receiver is more dependent on the antenna and its installation than any other single factor.

In the antenna installation consisting of the television antenna and the transmission line to the pre-amplifier there are four major items which should be considered in the design of a long range system,

1. antenna
2. location and height
3. transmission line
4. matching

These items cannot be thought of individually as each is dependent upon the others. To begin with, television transmission is a line-of-sight transmission or more exactly it has 1.156 times the line-of-sight range,

$$D = 1.42\left(\sqrt{H_T} + \sqrt{H_R}\right)$$

where D = television range in miles

H_T = transmitter antenna height in feet

H_R = receiver antenna height in feet

Because the attenuation of a r.h.f. signal along the line-of-sight path, with ranges up to many thousands of miles, does not materially effect signal reception the most important consideration in the antenna installation is height. There are however, some very practicable limits to this as can readily be seen from the following numerically example.

Assume the transmitting station is atop the Empire State Building at a height of 1500 ft. With the receiving antenna at the following heights, the television-range will be as follows,

Height in feet	Range in miles
10	58
50	63
100	68
150	70
1500	108

It is quickly obvious that this is not the solution to the long range problem, unless the receiver is located on a mountain peak, and the best we can do is place our antenna as high as practicable and in the best position with relation to reflections and interference from surrounding objects.

Since there is limit to the height of the receiving antenna, the next item is choice of the antenna. As much energy as possible must be taken from the atmosphere and this can be done only by a complex array.

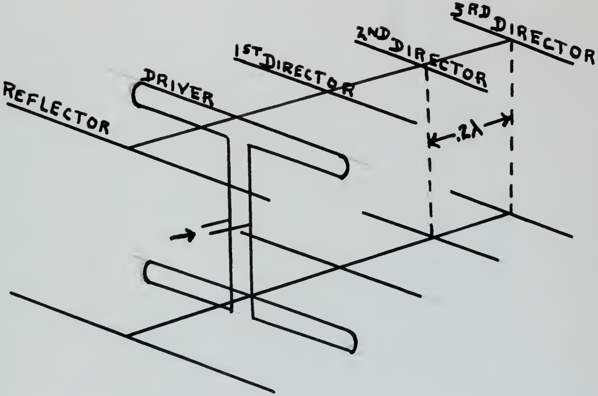
Several factors must be considered in the design of the array,

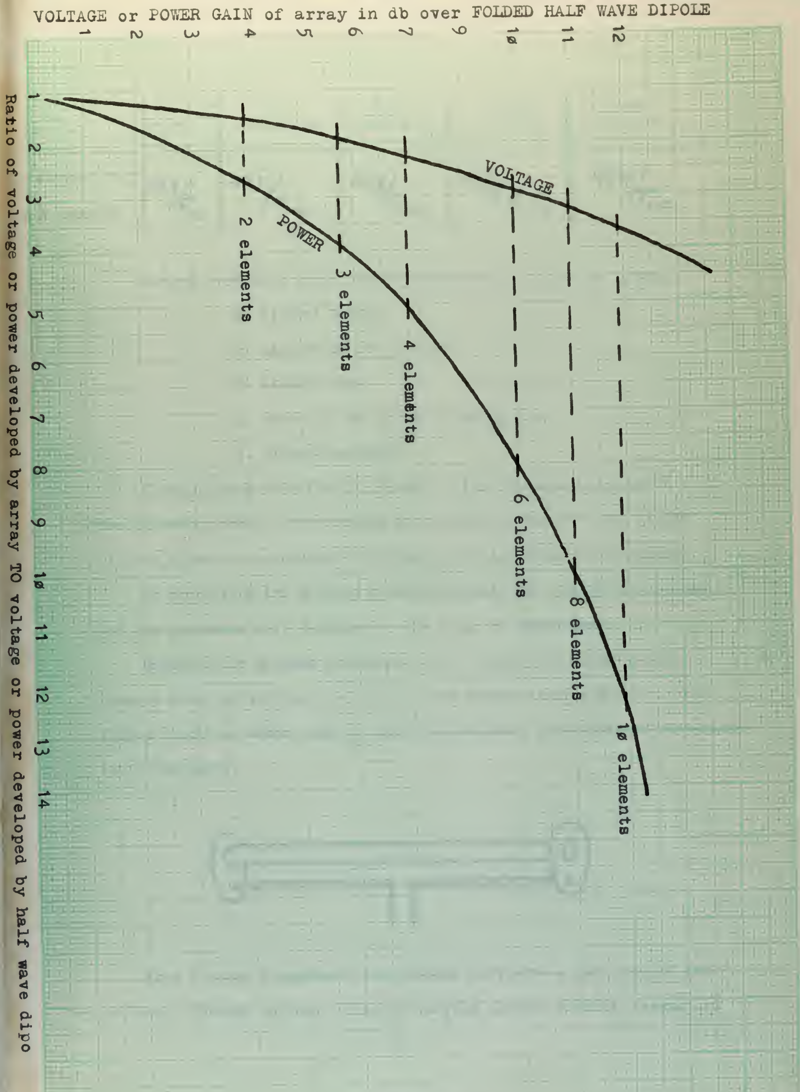
1. gain
2. band pass
3. size
4. complexity
5. matching
6. pattern
7. front-to-back ratio

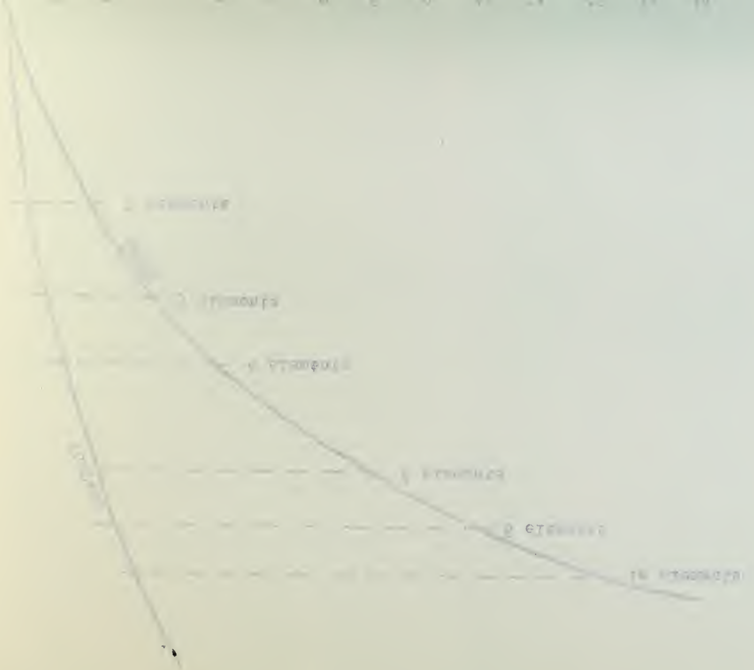
Technical information regarding antenna arrays is needlessly

confused by the maze of conflicting performance figures which is available. Given below is a chart giving an accurate estimate of the maximum gain available from antenna arrays with various numbers of half-wave elements as compared with a folded dipole. Both the reference folded dipole and the array in question are terminated in a 300 ohm load so that the result is a gain figure which takes into consideration not only the true gain of the antenna but also the mismatch loss which occurs into a 300 ohm load. The elements are spaced for maximum gain which is produced by a change in current distribution on the antenna elements.

The types of arrays which are available for use in an antenna system are very numerous. An example of a practical type as shown is the fine element stacked Yogi.




VOLTAGE or POWER GAIN of array in db over FOLDED HALF WAVE DIPOLE
1
2
3
4
5
6
7
9
1ø
11
12
Ratio of voltage or power developed by array TO voltage or power developed by half wave dipo
1
2
3
4
5
6
7
8
9
1ø
11
12
13
14
VOLTAGE
POWER
2 elements
3 elements
4 elements
6 elements
8 elements
1ø elements



	Driver	Reflector	1st Director	2nd Director	3rd Director
5 element Yogi .2λ spacing	462/F_{MC}	490/F_{MC}	442/F_{MC}	438/F_{MC}	434/F_{MC}

Several variables effect the characteristic of such an array,

1. element length
2. separation of elements
3. element size
4. ratio of the driver elements size
5. element material

A rough thumb rule for the length of the elements is given in the following table. For maximum gain further changes in the length of each element is necessary depending upon element size and spacing.

By decreasing the spacing between elements the gain will increase but the impedance will decrease to the order of 10-20 ohms.

Matching the antenna impedance to the transmission line can be brought about by varying the ratio of the driver element sizes of the folded dipole as shown, and by using transmission line with low characteristic impedance.

From a noise standpoint, the antenna must have a band pass of 6mc or less. Through antenna tuning by varying element spacing, length and

diameter, the Q of the array impedance may be made any desired value from 10 to 50 thereby controlling the bandwidth to quality ratio of the incoming signal.

While the Yogi antenna, from the standpoint of adjustment, is considered less complicated than other multi-element arrays, there are many variables envolved and the adjustment procedure is a long time-consuming operation.

CHAPTER X

CONCLUSION

In order to accomplish the reception of long range television signals each independent receiving installation must be undertaken as a separate problem.

First, a complete survey of surrounding television transmitting stations must be made with relation to such items as,

1. distance from transmitter to receiver
2. terrain separating transmitter and receiver
3. height of transmitting antenna
4. power output of transmitting station
5. frequency of transmitting station

The design of the long range antenna system should commence with the antenna installation. The practicability of locating a more or less complex antenna array as high as possible is considered. A factor which will enter in the positioning of the antenna will be that of propogation interference from local surroundings such as trees, building and power lines which might tend to shield the immediate vicinity of the proposed installation.

Since the system must of necessity be single channeled, a good choice from the standpoint of maximum gain, bandpass, simplicity of construction, installation and adjustment in comparison to other antenna arrays appears to be one which is made up of a pair of four or five element Yogi antennas. The feasibility of using a more complicated array or a rhombic should however, not be overlooked.

The system, as far as possible, should be matched throughout. Matching the antenna to the transmission line will bring about the transfer of the maximum amount of signal energy to the transmission line. Matching the transmission line to the pre-amplifier input will eliminate standing wave on the line and thereby minimize these losses and minimize reflections on the line.

Choice of transmission line is another problem which is first dependent upon the individual situation. If local disturbances and interference is present shielded twin lead or shielded coaxial cable may be necessary, while in areas relatively free of local interference normal twin lead with its low loss characteristic should be used. The transmission line characteristic impedance should be considered relative to both matching and the input circuit of the pre-amplifier.

For best results the basic low noise pre-amplifier circuits appear to be a choice between the Wallman amplifier or a pre-amplifier employing triodes operating in push pull followed by low noise pentode amplifier circuits. Actual pre-amplifier circuit design and component positioning should be given a high priority in the design of the overall system.

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